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by

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THESIS

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June 1969

A Nanosecond Microwave Burst Generator
Using Resonant Buildup For
Output Peak Power Enhancement

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ABSTRACT

A system has been developed for generating nanosecond-order pulses of microwave energy with the output pulse amplitude considerably higher than the input CW source level. The system uses a PIN diode switched waveguide cavity for a pulse forming device. The output pulse width is shown to be directly proportional to the actual cavity length. Minimum pulse width and degree of enhancement of the output amplitude are shown to be a function of the switching time and loss characteristics of the switch. Pulse widths as narrow as 4.3 nsec and enhancements as high as 4.8 dB were obtained using pulse repetition rates from zero to 500 KHz.

The output level enhancement feature of this system is its chief advantage over other nanosecond RF pulse generators. Other advantages of the system are its relative low cost, simplicity of tuning, versatility of pulse width and repetition rate, and its adaptability for rugged construction.

TABLE OF CONTENTS

LIST OF ILLUSTRATIONS

TABLE OF SYMBOLS AND ABBREVIATIONS

cm	centimeters
CW	continuous wave
dB	decibels
DC	direct current
f	frequency
KHz	kilohertz
LCR	inductance, capacitance, resistance circuit
ma	milliamperes
mw	milliwatts
nsec	nanoseconds
PRR	pulse repetition rate
Q	quality factor of resonant circuit
RF	radio frequency
SPST	single pole single throw
t	time
V	voltage
v	velocity
α	amplitude loss factor
Γ	reflection coefficient
λ	wavelength
π	3.14159......
τ	time

ACKNOWLEDGEMENTS

Sincere appreciation is extended to Professor George L. Sackman who acted as advisor on this project, and to Professor David B. Hoisington. Their valuable technical advice and encouragement was necessary and very much appreciated. Their high degree of professional competence and admirable personal bearing provided the motivation required by the author.

I must also thank my wife and family. The selfless understanding they showed for long periods of neglect is very much appreciated.

I. INTRODUCTION

Generation of nanosecond pulses of microwave energy has become the subject of much discussion in recent years due to the rapid advancement in high-speed microwave switching techniques. As early as 1957, Garver [1] demonstrated that switching speeds in the order of three nsec were possible using a germanium point-contact diode in a waveguide mount. Further work by Garver [2] resulted in several specific switching techniques. In a paper published in 1959 [3] Garver reported an observed switching speed of 1.5 nsec with the germanium diode. The past ten years has seen an acceleration in the progress of both waveguide and TEM mode diode switching techniques (see Refs. [4] through [10]) using a variety of solid-state junction and bulk devices. This progress in the state of the art has resulted in the wide commercial availability of microwave switches with switching times of a few nsec. Switching diodes are now available with picosecond-order switching times, and it can be expected that in the very near future these diodes will be available in microwave switches.

A large number of schemes have been developed which result in nsec microwave pulses (Refs. [5], [6], [7], [9], and [10] pertain). Most schemes involve one of two basic approaches to the problem. One common approach is to interrupt a CW source in a time sequenced switching scheme such as that of Young [9]. These methods are effective, but require very complicated switch driver systems. Dawson [6] describes an example of the other basic

method involving the application of large-amplitude nsec voltage pulses to certain special junction devices producing phase-locked bursts of microwave power. The method produces very narrow pulses but system tuning is difficult. Other similar systems such as those discussed by Shoemaker [11] have been used with varying degrees of success.

The object of this thesis was to investigate the feasibility of a system whereby a low-power CW microwave source could be used to produce high peak power pulses with nsec-order time duration. Such a system could make feasible the use of miniature CW sources (Gunn or Varactor types are commercially available) in pulsed applications with outputs in the order of watts. The system would find specific application as the pulse generator in Shoemaker's scheme [11] for finding and analyzing faults in waveguides.

The system developed for this thesis utilizes the concept of resonant build up in a waveguide cavity to provide an enhancement of the output pulse level over the CW source level (see Fig. 2). The critical element of the system is a high-speed microwave SPST switch.

Attempts were made to produce a switch similar to Garver's 1N263 waveguide mounted diode switch [2]. Silicon diodes were found to be too slow for this application. Since the only germanium diodes available were glass cartridge 1N263 diodes, cartridge impedance problems were insoluble (Refs. [12] and [13] pertain). Several methods of reflecting a short across the waveguide by placing a shorting diode in a tuned stub were investigated.

The above attempts were unsuccessful because the losses in the switch and the inability to obtain sufficient isolation characteristics lowered the Q of the cavity below a useable value.

A PIN diode SPST switch was purchased which provided the required isolation and sufficiently low loss characteristics to prove the feasibility of the system. Basically, a PIN diode switch consists of two or more PIN diodes shunted across an RF transmission line. The switching action is dependent upon the charge storage in the PIN junction. During the RF-ON state (reverse bias on diodes) the stored charge is negative (polarity) and the device appears as a capacitance shunted across the transmission line. Switching occurs when the diode is forward biased, building up a positive stored charge and providing a high conductance shunted across the transmission line. The junction conductance is approximately zero for negative charge storage. Otherwise, it is approximately equal to the charge stored in the diode.

II. THEORETICAL

In this section the general theoretical background is presented for the pulse generating system developed in this thesis. The basic concepts are presented, and theoretical calculations of output from the system are provided. The calculations use certain measured parameters of the system in order to provide comparison with experimental results.

A. GENERAL DISCUSSION

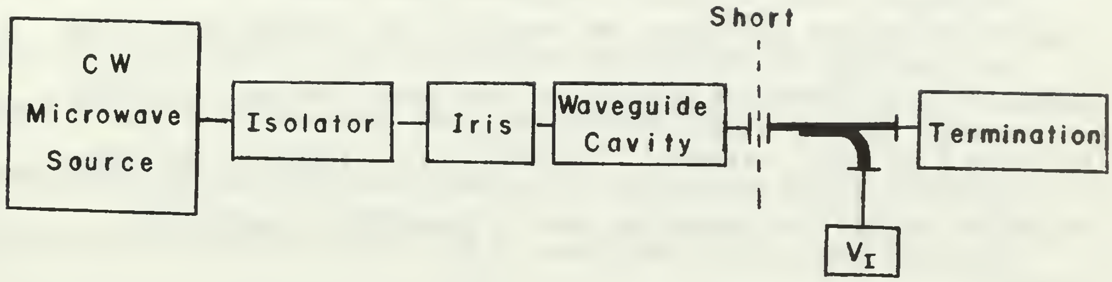
An iris-coupled short-circuited waveguide near resonance may be considered equivalent to a series LCR circuit shunted by a reactance (that of the iris). The unloaded Q of this circuit is given by Montgomery [14] to be

$$Q_u = \frac{\pi \lambda_g^2}{2\,(\alpha \ell)\,\lambda^2}\ , \qquad (1)$$

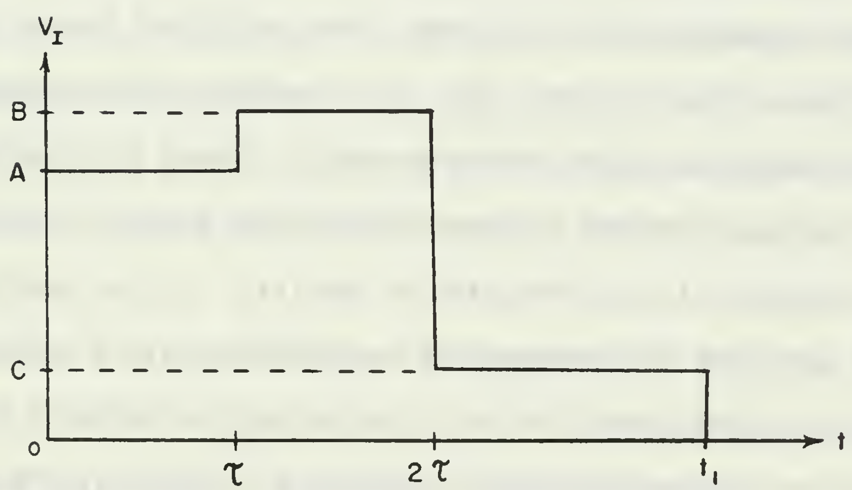
where α is the amplitude loss per unit length in the waveguide and ℓ is the length of the waveguide. If the waveguide is terminated in its characteristic impedance on the generator side of the iris, the loaded Q is related to unloaded Q by the relation

$$\frac{1}{Q_L} = \frac{1}{Q_u} + \frac{2}{\pi}\left(\frac{\lambda}{\lambda_g}\right)^2\left(\frac{1}{B}\right)^2\ , \qquad (2)$$

where Q_L is the loaded Q and Q_u is the unloaded Q. B is the normalized susceptance of the iris. If B is large, the second term in Eq. (2) may be neglected since $\left(\frac{1}{B}\right)^2$ is very much less than unity. Hence, $Q_L \approx Q_u$ in this case.



(a) Pulse Generator Circuit.



(b) Idealized Output Pulse Shape.

Figure 1.
Pulse Generator Circuit

Inside the waveguide cavity a standing wave exists in which the E-field maximums are Q_L times the E-field maximums in the wave incident on the coupling iris. This is analogous to the case in the series LCR where the magnitude of voltage across the inductor is Q_L times the applied voltage magnitude as discussed in Cruft [15], for instance. The effect is a resonant build-up of energy within the waveguide cavity depending in magnitude upon Q_L.

The waveguide cavity can be "dumped" by suddenly removing the short and connecting the cavity to a waveguide (see Fig. 1.(a)). This provides a "square" pulse of carrier, analogous to discharging a section of transmission line into a load of equal characteristic impedance (a normal resonant cavity suddenly dumped, produces the decaying exponential pulse obtained by discharging a lumped capacitance into a load). Here a wave is initiated which propagates down the waveguide toward the termination. A wave of equal amplitude (6 dB down from the resonant level in the cavity) is initiated which propagates in the direction of the iris. At the iris the reverse wave sees a high impedance and reflects with a positive reflection coefficient. The result is an initial pulse of RF energy (levels A and B of Fig. 1.(b)) which is physically twice the length of the waveguide cavity. This is followed by a lower level wave which represents the transmission from the CW source through the iris (level C in Fig. 1.(b)). Let τ represent the time required for the wave to propagate the length of the cavity. If the short is replaced at the exact time the pulse passes the plane of the short ($t_1 = 2\tau$), the lower level "tail" (level C) will be cut off.

Level B, occurring during $\tau \leq t \leq 2\tau$, is due to the in-phase summation of that amount of generator energy transmitted through the iris and the reverse wave reflected from the iris. Depending upon the magnitude of the reflection coefficient (Γ) at the iris level B can be either less than, or greater than level A. The second term of Eq. (2) corresponds to the inverse of the external (or radiated) quality factor (Q_E), and since this term is almost zero (as shown above) the Γ of the iris is expected to be high. Since

$$Q_L = \frac{1}{1/Q_u + 1/Q_E} = 2\pi \frac{\text{Energy stored in cavity}}{\text{Energy dissipated/cycle}} ,$$

a high Q_E implies more energy stored in the cavity and less energy radiated out of the cavity through the iris. This implies a high magnitude of Γ at the iris. Thus, level B is expected to be higher than level A as shown in Fig. 1.(b). The resulting pulse is directly proportional to the length of the waveguide cavity and is of greater peak amplitude than the incident CW wave by a factor which is a function of Q_L.

The above discussion and Fig. 1.(b) assumes that the short is removed and reinserted in zero time. In the actual circuit a microwave PIN diode switch with a switching time of three nsec is used.

B. THEORETICAL CALCULATIONS

In this subsection the relation between length of the waveguide cavity and the observed time duration of the output pulse will be developed. The maximum resonant build up and predicted output pulse amplitude will also be calculated from theory.

1. Pulse Width as a Function of Cavity Length

The length of waveguide cavity required for a one nsec pulse can be determined by considering the distance a wave would travel at the group velocity (v_g) in one nsec. It is known that

$$v_g = \left(\frac{\lambda}{\lambda_g}\right) c \qquad ,$$

where c is the velocity of light in free space. Also,

$$\lambda_g = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{\lambda_c}\right)^2}}$$

where λ_c is the cutoff wavelength and λ_g is the phase (guide) wavelength. As explained previously, the trailing edge of the pulse travels the length of the cavity twice (from the switch end to the iris and back). In other words, the effective length in space of the pulse train emerging from the cavity is twice the physical length. Since the wave travels v_g meters per second, the resulting pulse should be one nsec long for each $\frac{v_g}{2} \times 10^{-9}$ of actual cavity length.

2. Output Pulse Amplitude

The resonant build up in the cavity and the amplitude of the output pulse can be determined from the Q_L of the cavity. Eq. (1) was previously shown to hold as an expression for Q_L under the condition of high iris susceptance. It should be noted that the loss coefficient (α) factor in Eq. (1) is considered to be the loss per meter of the waveguide (skin loss). When a PIN diode switch is used instead of a shorting plate the switch is found to have finite losses (as discussed later). For purposes of this theoretical

computation the loss due to the switch (cavity end loss) is considered to be additive with the skin loss in the waveguide. In other words, the end loss can be accounted for as if it were distributed uniformly along the waveguide. Q_L, which was shown to be equal to Q_u in this case, is modified from Eq. (1) for this application as follows:

$$Q_L = \frac{\pi \lambda_g^2}{2(\alpha' \ell) \lambda^2} \tag{3}$$

where α' is the combined skin loss and cavity end loss. Using Eq. (3) and parameter values calculated from experimental data, a Q_L can be determined for a cavity of given length.

With the switch acting as a short at the end, the cavity may be considered as a simple one-port device. A normalized voltage (V) term proportional to the E-field will be used here as defined in Montgomery [14]. V_I represents the "voltage" applied to the terminals of the one-port device (cavity) and V_C represents the "voltage" inside the cavity. As was previously stated, the RLC circuit analogy holds near resonance and it can be shown that

$$V_C = Q_L V_I \quad , \tag{4}$$

or

$$Q_L = \frac{V_C}{V_I} \quad . \tag{5}$$

It is now clear that 20 log (Q_L) represents the resonant build-up in the cavity in dB above the incident wave amplitude.

The skin loss is very small compared to the end loss when using the PIN diode switch, hence the Q_L is primarily a function of end loss and is relatively independent of the length of wave-

guide cavity. Assuming that $\alpha \ll \alpha''$, Eq. (3) becomes

$$Q_L = \frac{\pi \lambda_g^2}{2 \alpha'' \lambda^2} , \qquad (6)$$

where α'' is the cavity end loss due to the PIN diode switch.

The value for Q_L determined using typical PIN diode specifications ($Q_L \approx 4.3$) at first appears to be extremely low compared to values in the thousands normally found for resonant cavities. However, when it is considered that the end loss diminishes the circulating wave on each pass (by 2 dB in this case), the low value of Q_L is understandable. This is verified by reference to Eq. (6) where it is easily seen that the factor $\frac{\pi \lambda_g^2}{2 \lambda^2}$ is of the order of unity. Thus, for very large values of Q the α'' factor (expressed as a fraction) must be much less than unity.

Although a resonant build up is expected within the cavity, we cannot expect the output pulse peak amplitude to be at this level. As was mentioned previously, the initial pulse amplitude is 6 dB below the cavity level due to the initiation of a reverse wave. There is an additional loss when the output wave passes through the switch assembly. This loss is the insertion loss of the switch. Hence, from this we may determine the relative peak amplitude of the output pulse using the values measured:

frequency (f) = 8.97 GHz,

wavelength (λ) = 3.34 cm,

skin loss (α_{dB}) = 2×10^{-3} dB/cm,

cavity end loss (α''_{dB}) = 2 dB,

switch insertion loss = 1.1 dB,

Relative peak amplitude = 12.7 dB - 6 dB - 1.1 dB = 5.6 dB.

Thus it is seen that with the simple circuit of Fig. 1.(a) operating with a high-speed SPST microwave (PIN diode) switch in place of the short, it is possible to obtain very narrow (nanosecond-order) pulses of microwave energy. The pulse width is controlled by simply varying the length of waveguide cavity. With increased length of cavity it can be shown that Q will increase only slightly. Thus, a slightly greater enhancement over the CW source level can be expected for added cavity length, but with increased pulse width. Hence, in addition to the pulse generation, this system has the capability of an appreciable enhancement of the output pulse amplitude over the CW input level. This enhancement is strongly dependent upon switch losses in the "RF-OFF" condition.

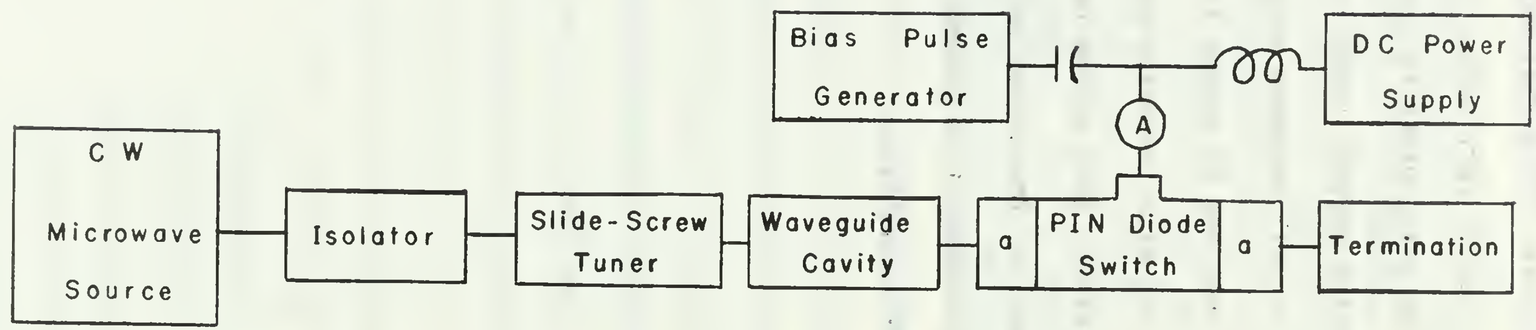


Figure 2.
Experimental Circuit.
Item "a" is a combination of a Hewlett Packard X281A waveguide-to-coax adapter with a UG-27C/U and a UG-57B/U connected.

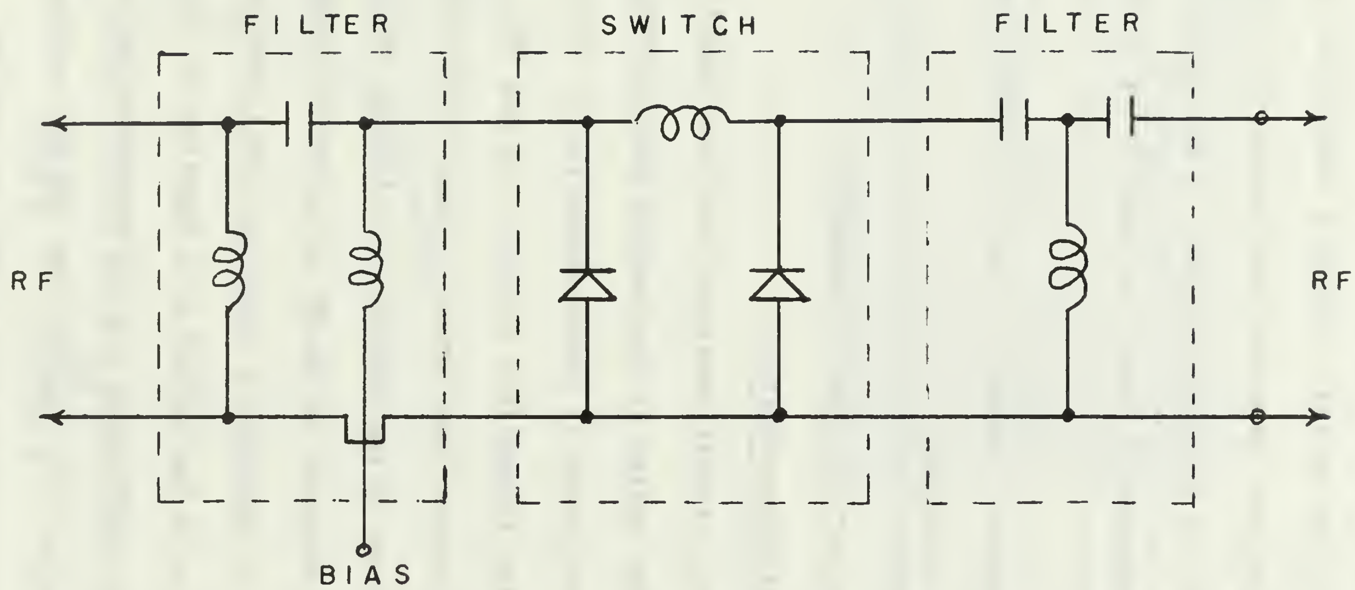


Figure 3.
SPST Microwave PIN Diode Switch

III. EXPERIMENTAL RESULTS

This section presents the actual circuit developed for the thesis and discusses certain important components of the system. Experimental data are summarized and compared to the theoretical calculations of the previous section. A discussion of the results obtained by attempting to use a narrower drive pulse to the switch is presented.

Fig. 9. presents a summary of the experimental results for the parameters of interest.

A. DISCUSSION OF TEST CIRCUIT COMPONENTS

1. SPST Microwave Switch

The test circuit developed for this report is shown in Fig. 2. The heart of the system is the Hewlett Packard HP-33571 SPST microwave switch which is shown schematically in Fig. 3. As was indicated in the theoretical section, the end loss due to the switch is the controlling factor in determining the magnitude of resonant build up in the cavity. The loss in the switch is due to the filter networks at the RF ports and to diode losses in the switch. Fig. 4. and Fig. 5. show the cavity end loss and isolation characteristics of the switch as a function of forward bias current. The reflection coefficient (Γ) and insertion loss were found to be almost constant for reverse bias (RF-ON) voltages from zero to -30v (maximum reverse bias = -36v) as follows:

$$\Gamma \;=\; 0.13 \quad (20 \log \Gamma = -17.5 \text{ dB}),$$

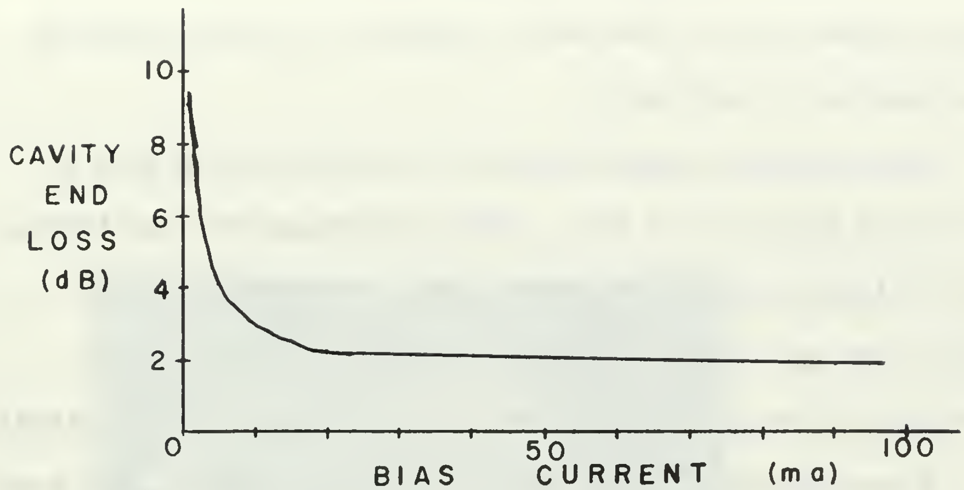


Figure 4.
Cavity End Loss vs. Switch Bias Current

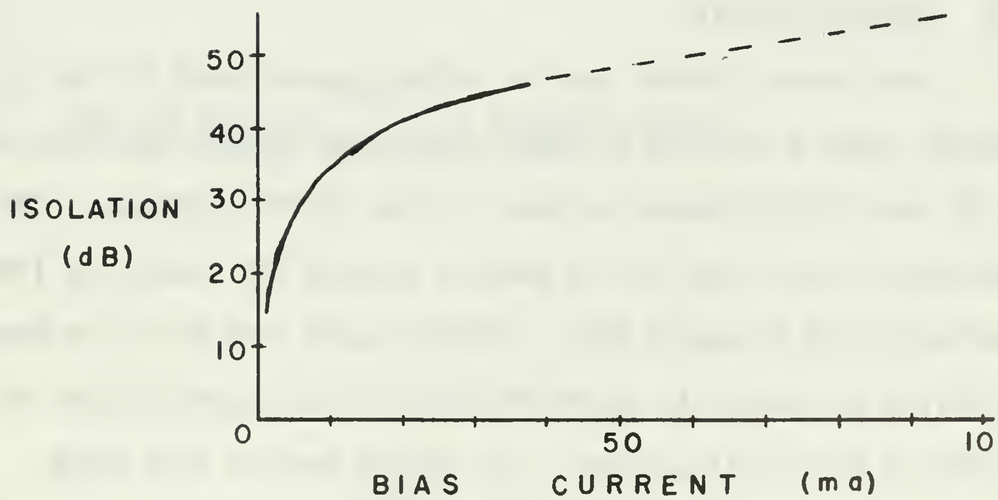


Figure 5.
Isolation vs. Switch Bias Current

Insertion loss = 1.1 dB.

The above values are for the switch assembly including waveguide-to-coax adaptors on both ends.

Manufacturer's specifications of switching time gave a typical RF-ON time of nine nsec. Test data showed this switching time to be three nsec for the switch used (see Fig. 6.). The switching time was found to be somewhat sensitive to the rise time of the biasing pulse (this is discussed in more detail later).

A very important advantage of the switch used is that properly biased it has no low limit on pulse repetition rate. The system can deliver nsec pulses at any desired repetition rate up to a limit controlled by the switching diode recombination time [8] and the bias pulse width. The diode recombination time is of the order of picoseconds. The test circuit was operated at repetition rates from a few hertz up to 500 KHz with no noticeable change in output RF pulse characteristics.

2. Biasing Circuit

The biasing circuit for the switch is indicated in Fig. 2. A Hewlett Packard HP-6216A DC power supply was used to provide a positive DC bias (90 milliamps) current for the RF-OFF condition. The inductance in this bias line is used to prevent the bias pulse from grounding in the DC supply [8]. The bias pulse from the pulse generator drives the switch to the RF-ON condition by overriding the DC bias with a 17v negative pulse. The leading edge of this pulse initiates the switch RF-ON condition which dumps the waveguide cavity. The pulse generator used provided a minimum 25 nsec pulse width with rise and fall times adjustable upward from 10 nsec.

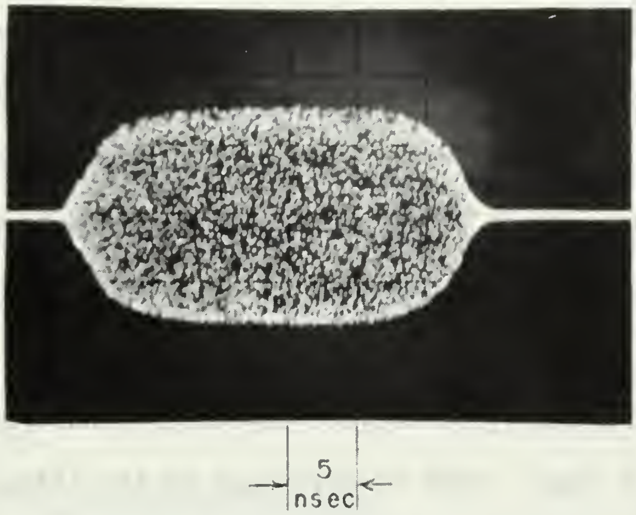


Figure 6.
8.97 GHz Switched with the HP-33571 Switch.
Switch-ON Time = 3 nsec.

3. <u>Coupling Iris</u>

Several different coupling methods were investigated for obtaining an iris-coupled short-circuited waveguide. A thin conducting plate type iris was employed at first, but was found to be too difficult to tune. An E-H tuner was employed and found to work very well except that it was not possible to return to an exact setting after detuning (an E-H tuner with micrometer arms was not available). A slide-screw tuner was finally decided upon because it is easily adjusted over a wide susceptance range and can be returned to a previous setting with precision. One disadvantage of the slide-screw tuner is that the probe introduces a small ohmic loss as well as a susceptance in the line. The actual normalized conductance was found to be of the order of 0.6 mhos. This disadvantage of the slide-screw tuner was accepted as tolerable considering the advantage it provided for the experimental procedure. Degradation of circuit performance using the slide-screw tuner was negligible.

One factor which must be considered in this application is the distance from the probe of the slide-screw tuner to the cavity waveguide flange. This length is actually part of the cavity and must be considered when examining output pulse width compared to cavity length.

4. <u>Microwave Source</u>

The source used for experimental data was the Hewlett Packard HP-670HM klystron generator with the HP-717A power supply. Output power for the experiment was about 30 mw. The generator

power level was found to have negligible effect on the results. The only limitation to the generator power level is the switch, which allows a maximum incident power of one watt.

5. Oscilloscopes

Sampling oscilloscopes were used for viewing the output waveforms. Photographs of video (detected RF) waveforms included in this report were taken on the Hewlett Packard HP-185B sample scope. Photographs of RF pulse envelopes used here were taken on the HP-140A scope using ancillary sampling components as required. It was found that viewing the actual RF pulse very much simplified the interpretation and improved the accuracy of data, as is indicated in the literature [8].

6. Sample Waveguide Cavities

Four different lengths of cavity were examined in this thesis. The lengths of waveguides used as cavities here were not specifically selected. On-hand waveguide sections (copper waveguide) were cleaned before using. A dimethyl polysiloxane release agent followed by several alcohol rinses provided sufficient cleaning.

The actual lengths of waveguide used and their cavity lengths are listed below:

Sample	Waveguide (cm)	Cavity Length (cm)
1.	37.5	49.44
2.	45.3	56.79
3.	75.5	85.42
4.	120.8	132.74

I Sample No.	II Cavity Length (cm)	III Pulse Width (nsec)		IV Cavity cm / Pulse nsec		V Peak Enhancement Over Input (dB)	
		Meas	Calc	Meas	Calc	Meas	Calc
1.	49.44	4.5	4.37	11.0	11.3	4.3	5.6
2.	56.79	5.0	5.03	11.3	11.3	4.5	5.6
3.	85.42	7.5	7.56	11.4	11.3	4.7	5.6
4.	132.74	11.5	11.7	11.5	11.3	4.8	5.6

Table 1.
Summary of Measured and Calculated Results.

The physical lengths of the waveguides differ from the cavity lengths in each case because of the additional length in the slide-screw tuner and the apparent cavity length resulting from the switch assembly. The additional length in the slide-screw tuner depends upon the optimum position of the probe, and is different for each waveguide. The apparent length of the switch assembly results from the delay incurred in the waveguide-to-coax transitions and the filter section of the switch (see Fig. 3.). This value is a constant at a given frequency. All future reference to the waveguide cavity sections will be by sample number (listed above). Computations were made considering the cavity length.

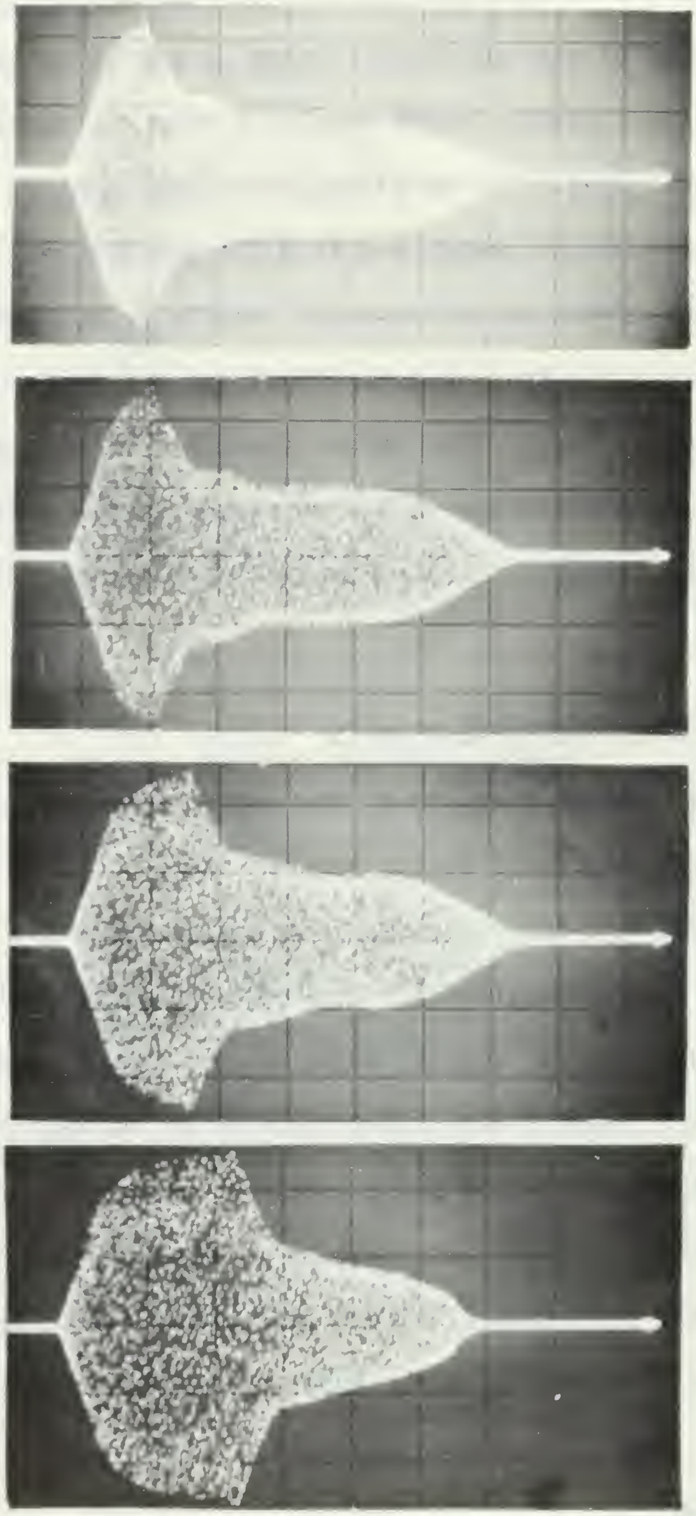
B. PRESENTATION OF EXPERIMENTAL RESULTS

Figure 7. shows the RF envelope of the output pulse obtained for each of the four waveguide cavity samples tested. Table 1. summarizes the pertinent data obtained from these waveforms.

The length of cavity required for each nsec of output pulse was calculated to be 11.3 cm per nsec of pulse width. The comparisons presented in Table 1. show a very close correlation between theory and data. The measured value of pulse width in column III of Table 1. was obtained directly from the sample-scope presentation of the output pulse envelope. The differences between measured and calculated values in Table 1. are well within the error to be expected due to scope interpretation.

1. Pulse Shape

The shape of the output pulses also correlates well with the theoretical model. The following discussion holds for each of the samples



(a)
Sample No. 1.

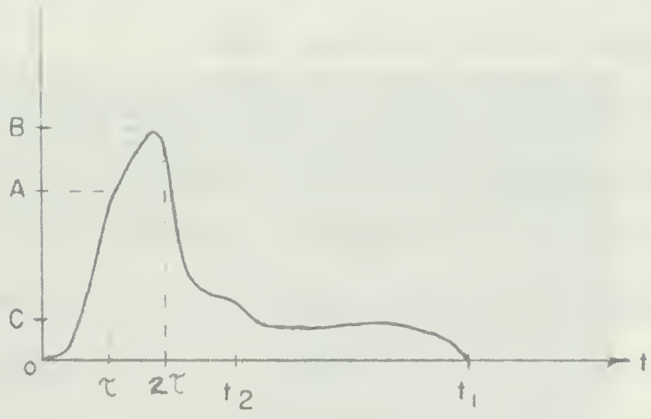
(b)
Sample No. 2.

(c)
Sample No. 3.

(d)
Sample No. 4.

Figure 7.
RF Output From Experimental Circuits.
Horizontal Scale = 5 nsec/Division

(a) Detected RF Envelope Using Sample No. 2.
Horizontal Scale = 5 nsec/Division



(b) Mode Output Pulse Shape.

Figure 8.
Output Pulse Shape.

so that Fig. 8.(b) can be considered as typical for purposes of explanation. Direct comparison of time and amplitude references with Fig. 1.(b) shows that the theoretical calculations were quite close to measured values, the only difference being the leveling off at $2\tau<t<t_2$. This phenomenon appeared in each of the samples but it's time duration is not a function of sample length.

The explanation for the above mentioned leveling off is based on the fact that the switch cannot change state in zero time. At the first instant the initial wave is totally reflected from the switch. Then, as the switch changes state the amount of reflection decreases until the switch is in the fully "ON" condition. The effect is an additional reflected pulse which follows the trailing edge of the previously discussed reflected wave (represented by $\tau<t<2\tau$ in Fig. 1.(b) and Fig. 8.(b)). This pulse is approximately the switch-on time of the switch in duration, and hence is not a function of cavity length. Close inspection of the time duration of the phenomenon shows that in fact it is approximately equal to the switching time.

2. Pulse Enhancement Over CW Input

Reference to Table 1. column V shows the close correlation of theoretical and measured enhancement. The theoretical calculation of 5.6 dB enhancement is, however, somewhat higher than the measured value. The difference is due to the fact that the finite switch-on time causes some energy to be reflected initially (during cavity dump). This energy is in effect subtracted from the initial pulse and added back in at time $t = 2\tau$ (in Fig. 8.(b)) after the initial pulse peak has passed. Thus, the deviation from theoretically predicted peak enhancement is due to the switching action and is a function of switching time.

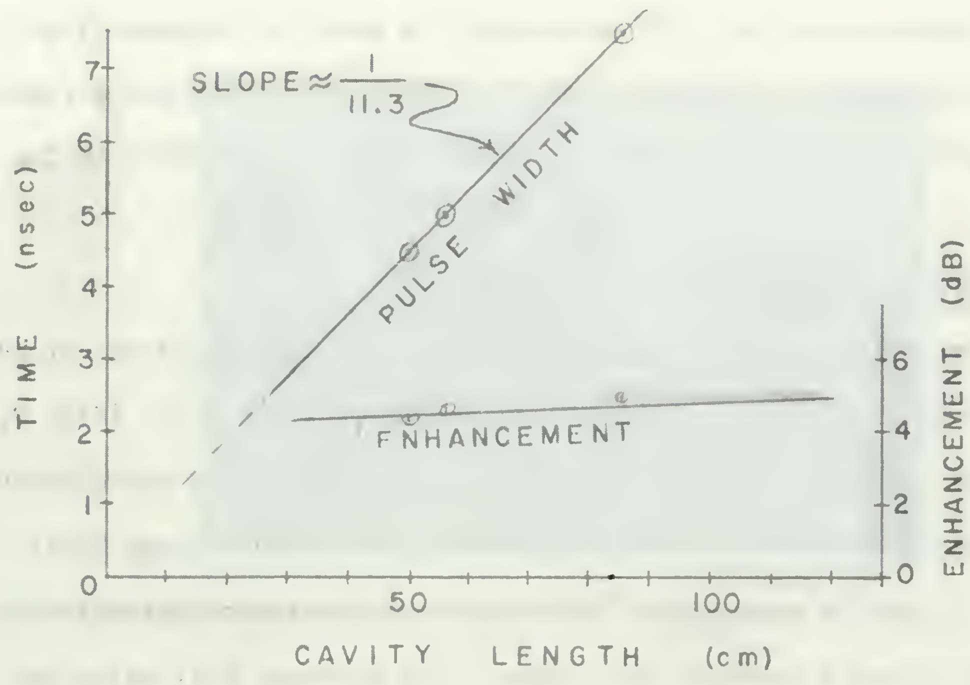


Figure 9.
Graphic Summary of Results.

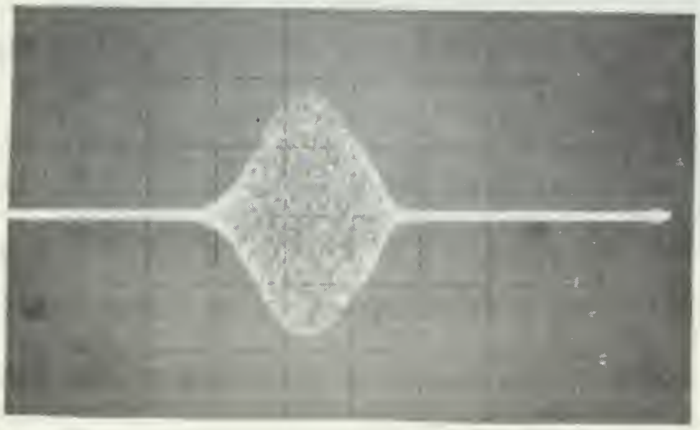


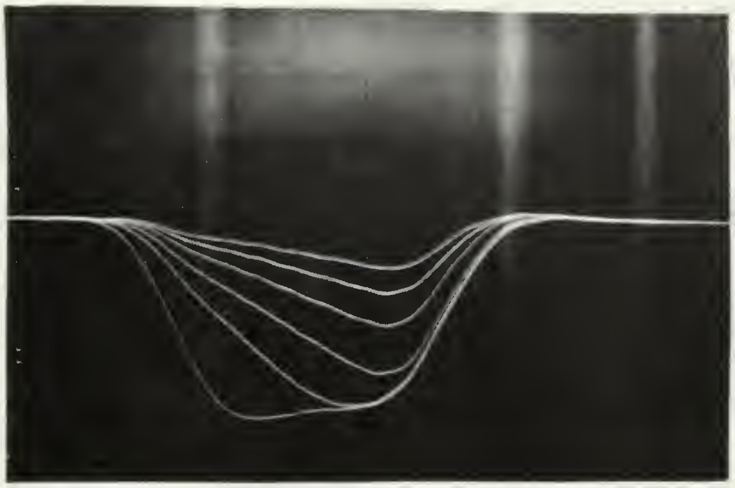
Figure 10.
RF Pulse Envelope with Reduced Bias Pulse Width.

The measured value of enhancement is seen to increase slightly as a function of cavity length (see Fig. 9). This effect was a result of an increase in Q with increase of volume to end-loss ratio in the cavity.

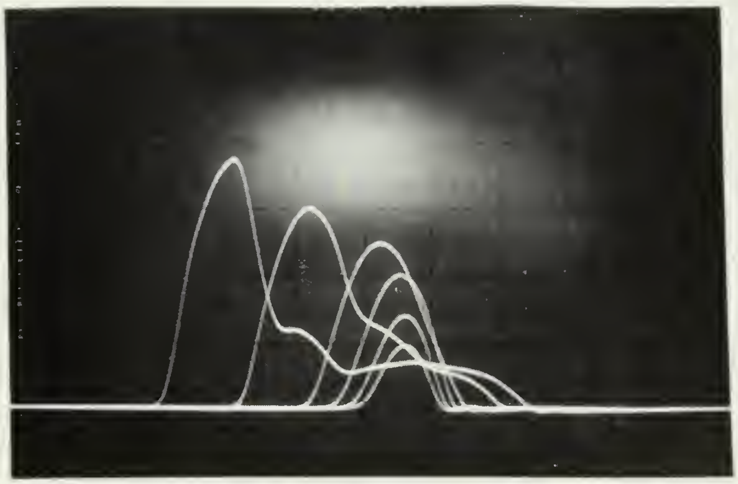
3. <u>Effect of Narrower Switch Drive Pulse</u>

It seems logical to assume that if the switch could be returned to the RF-OFF (short across waveguide) condition at $t = 2\tau$ (Fig. 8.(b)) the output pulse would have a more desirable shape. This would cut off the switching transient "plateau" and the source feed-through "tail." Fig. 10. is the RF envelope of the output pulse obtained using sample cavity No. 2 and a narrower bias pulse. The narrower bias pulse was obtained by increasing the rise time of the 25 nsec (narrowest available) bias pulse. According to Hewlett Packard application notes [8] switching time is increased by the increased rise time of the bias pulse. The "plateau" effect of the switching transient is therefore increased, which results in a serious degradation of RF output pulse amplitude, for the reasons previously discussed. Although the base of the bias pulse was unchanged, the switch on-time was decreased and the desired pulse shape was obtained.

Figure 11.(a) shows the effect of increasing the rise time of the bias pulse. Fig. 11.(b) shows the resulting effect on the output RF pulse shape. It should be noted that the decrease in output pulse amplitude is not a direct result of the decreased amplitude of the bias pulse at longer rise-times. One might be led to this false conclusion at first because of the superficial correlation of circumstances (i.e., lower amplitude bias pulse leading to increased switch insertion loss).



(a) Effect of Increasing Rise Time of the Biasing Pulse.



(b) Resulting Effect on Output Pulse.

Figure 11.
Effect of Increasing Bias Pulse Rise Time.
Horizontal Scale = 5 nsec/Division

However, as previously mentioned, the switch insertion loss is virtually constant from zero to maximum "ON" bias which negates this conclusion. A lower amplitude bias pulse does, however, decrease switching time.

Hence, it is seen that a clean (greater than 40 dB isolation from the source) burst of RF energy of a few nsec duration could be obtained from this system with the proper bias pulse applied to the switch. Further, the peak amplitude of the RF above the source level is limited primarily by the effect of switch losses on the attainable Q_L' of the cavity.

IV. CONCLUSIONS AND RECOMMENDATIONS

A. CONCLUSIONS

Until very recently the state of the art in the area of high-speed microwave switching was extremely limited. With the recent commercial availability of low-loss nsec switching devices it is possible to achieve an enhancement of the peak output pulse level over the CW input. Herein lies the chief advantage of this concept over other short microwave pulse generation methods presently available. Microwave pulses at X-band as short as 4.3 nsec with peak pulse levels as high as 4.8 dB above the CW source level were produced by the circuit. Pulse repetition rates (PRR) were varied from zero to over 500 KHz. Further, it was shown that the RF pulse width is a function of the waveguide cavity length and could feasibly be made even shorter. Much greater enhancement in output level is possible if lower loss in the switch (cavity end loss) and faster switching time were available.

The scheme has practical applications in areas of current interest in the microwave field. A recent thesis at the Naval Postgraduate School, Monterey, California [11] described the use of a narrow pulse of microwave energy in a time domain reflectometry application in waveguide, which would benefit by the compactness and rugged construction as well as the simplicity of tuning and circuitry made possible with this system. High resolution short-range radars [17] currently using more complicated pulse generation systems could be greatly simplified by this system. Use of a miniaturized solid-state microwave source and a mechanically tunable iris along with a wide-band switch such as the HP-33571 (or

similar models presently available from several manufacturers) in this application makes possible a relatively low-cost, rugged, stable, versatile (variable PRR and pulse width) pulsed microwave source as a test equipment.

B. RECOMMENDATIONS

The results obtained from experimentation with the system developed in this study suggests that further effort in this area is worthwhile. As has been previously stated, the microwave switching device is presently the main limitation of the system capability. Switching diodes which will provide picosecond switching times are currently available which might be applied in this area. Devices which were not obtainable for this thesis (due to lack of funds) are presently available which will provide one nsec switching time with much lower loss than the switch used [18].

The switch used in this investigation is capable of operation over a frequency band from one GHz to 12.4 GHz. This study was limited to the X-band case because of available time. Appropriate CW sources and hardware are available to extend this work to higher and lower frequency ranges.

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13. ABSTRACT

A system has been developed for generating nanosecond-order pulses of microwave energy with the output pulse amplitude considerably higher than the input CW source level. The system uses a PIN diode switched waveguide cavity for a pulse forming device. The output pulse width is shown to be directly proportional to the actual cavity length. Minimum pulse width and degree of enhancement of the output amplitude are shown to be a function of the switching time and loss characteristics of the switch. Pulse widths as narrow as 4.3 nsec and enhancements as high as 4.8 dB were obtained using pulse repetition rates from zero to 500 KHz.

The output level enhancement feature of this system is its chief advantage over other nanosecond RF pulse generators. Other advantages of the system are its relative low cost, simplicity of tuning, versatility of pulse width and repetition rate, and its adaptability for rugged construction.

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